Two-Stage Low-Noise Preamplifiers for the Amateur Bands from 24 cm to 12 cm

by J. Grimm, DJ 6 PI

Microwave converters for the amateur bands in excess of 70 cm usually operate with diode mixers such as types 1 N 21 – 1 N 23 or hp 2800 – hp 2817 etc. According to construction and extent of the circuit (image frequency trap !), the noise figures will be in the order of 8 to 16 dB. VHF COMMUNICATIONS has published preamplifiers in the past equipped with transistors AF 139, AF 279, BFR 34, which exhibited noise figures in the order of 5 dB. Bipolar SHF-stripline transistors are now available that allow noise figures in the order of 2 dB to be realized. Such transistors are manufactured by NEC, HP, Microwave Associates, Motorola and some others, and are available at prices that are still acceptable to the radio amateur. The following article is to describe two-stage amplifiers equipped with NEC-transistors of type NE 645 35 and NE 578 35.

Such low-noise preamplifiers increase the signal-to-noise ratio by the same factor as when increasing the output power by four to ten times. This increase of transmit power would cost many times that of such a preamplifier.

These preamplifiers represent a further development of similar preamplifiers described in (1), and (2). The amplifiers are constructed in stripline technology on a printed circuit board since this is far easier to calculate and construct than coaxial amplifiers using cavities. The losses in the PC-board material are only in the order of a fraction of a dB and are therefore most certainly acceptable when considering the simple and reliable construction that results from this.

For those readers not interested in the theory and calculation of such amplifiers, it is now possible for them to jump the following sections and to continue with the practical information that commences with Section 4.

1. CALCULATION OF THE MATCHING NETWORKS

1.1. Transistor Impedances

In order to calculate the matching networks, one must know the input and output impedances of the transistor at the required frequency, and for the various operating points. The manufacturers provide the complex s-parameters of the transistors for this purpose. Since the determination of the transistor impedances from the s-parameters has not been described in detail in amateur literature, this is to be discussed in more detail here.

Two methods of calculation are to be used in the following. All amplifier versions have been calculated according to these methods, and subsequently constructed. The differences in the measured values are given in the table in Section 5.

The s-parameters are determined by the manufacturer by terminating the input or output of the transistor with 50 Ω. The parameters s_{11} and s_{22} are the input and output reflection coefficients for this mode. Parameters s_{12} and s_{21} are also a measure how great an impedance variation at the input will have an effect on the output impedance, and vice versa.

Table 1:
An example of the s-parameters of transistor NE 578 35

U_{CE} = 8 V, I_C = 10 mA.
1st Column: Amount, 2nd Column: Angle

Frequency	s_{11}		s_{21}		s_{12}		s_{22}	
100 MHz	0,65	− 32	19,20	152	0,02	72	0,90	− 15
200 MHz	0,55	− 63	15,20	138	0,03	63	0,78	− 24
500 MHz	0,40	−127	9,80	102	0,05	52	0,50	− 38
1 GHz	0,40	−160	7,70	82	0,07	52	0,35	− 33
2 GHz	0,50	158	3,00	57	0,11	52	0,24	− 60
4 GHz	0,63	128	1,34	18	0,16	33	0,22	−128

The s-parameters are given in the data sheets for various frequencies and operating points in the form of a table (and sometimes also in the form of Smith diagrams); since these are complex magnitudes (vectors), they are given with amount and angle.

1.1.1. Simplified Calculation Method

In practice, the input and output of the transistor is not terminated with 50 Ω. If a slight mismatch is acceptable, one can save a considerable amount of calculation when one only uses the input and output reflection coefficients s_{11} and s_{22} for the calculation of the transistor input and output impedance.

The forward and return transfer coefficients s_{21} and s_{12} are not taken into consideration.

The impedances are calculated from the reflection coefficients using the following equation:

$$Z_{in} = Z_0 \times \frac{1 + s_{11}}{1 - s_{11}} \qquad (1)$$

where Z_0 = 50 Ω.

Since the reflection coefficients are complex magnitudes, the real and reactive impedance is calculated using the following equation:

$$Z_{in} = \frac{(1 - |s_{11}|^2) \times 50}{1 + |s_{11}|^2 - 2|s_{11}| \times \cos \sphericalangle s_{11}} + j\,\frac{(2|s_{11}| \times \sin \sphericalangle s_{11}) \times 50}{1 + |s_{11}|^2 - 2|s_{11}| \times \cos \sphericalangle s_{11}} \qquad (2)$$

The output impedance Z_{out} is also calculated in the same manner from parameter s_{22}, or from $|s_{22}|$ and $\sphericalangle s_{22}$. Example from table 1:

$$Z_{in} \text{ at 2 GHz} = 17.2 + j\,8.6\ \Omega$$
$$Z_{out} \text{ at 2 GHz} = 57.6 - j\,25\ \Omega$$

For those readers that do not wish to calculate the input and output impedances from this equation, it is possible for them to be taken in an approximate manner direct from the Smith diagram. Unfortunately, these diagrams are usually so small in the data sheets so that it is only possible to extract the impedances with a certain degree of inaccuracy. It is easier, and more accurate, when the tabular values are firstly transferred into a large Smith diagram. It can then be of any required size. This is achieved by firstly drawing the angles as lines from the center point of the diagram. The positive angles are to be found in the upper, the negative angles in the lower half of the diagram. The lengths of the lines result from the radius of the Smith diagram used, multiplied by the magnitude factor from the table.

Coefficient data is referred to an impedance of 50 Ω, which means that the values taken from a standardized diagram should be multiplied by 50 Ω.

Figure 1 shows an example of a Smith diagram in which all values have already been multiplied by 50 Ω. It contains the s_{11} and s_{22} values for a common emitter circuit which has been taken from Table 1. The

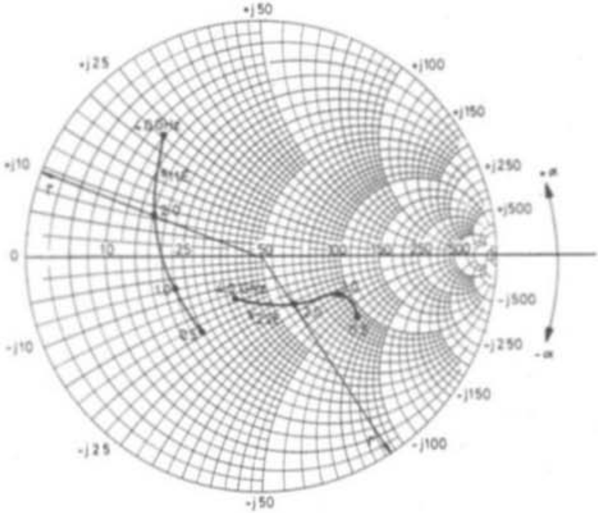


Fig. 1: Smith diagram with the input and output reflection coefficients of transistor NE 578 35 at U_{CE} = 8 V and I_C = 10 mA

lines have been drawn for a frequency of 2 GHz: s_{11} = 158° and 0.5 of the radius of the diagram, and s_{22} = −60° and 0.24 for the radius.

The following impedances can be read off:
Z_{in} approx. 17 + j 9 Ω;
Z_{out} approx. 60 – j 26 Ω.

1.1.2. Exact Method of Calculation

This method takes the reaction within the transistor and the actual terminating impedances of the transistor input and output into consideration. One requires the four s-parameters of the transistor at the required frequency, and the required operating point for calculation. This can be obtained with the aid of an s-parameter meter such as the hp 8410, which is hardly accessible to radio amateurs. However, the previously mentioned tables are available from the semiconductor manufacturers in which the four s-parameters are given for the various operating points in a frequency spacing of 200 MHz. The values for any required frequency can be obtained by inter or extrapolation.

The calculation of the corrected coefficients s_{11}^* and s_{22}^* for simultaneous matching of the transistor at the input and output (and thus the corrected impedances) would, however, be too extensive for this article. A calculation program for the TI 59 was developed, and further details regarding this are to be given in Section 6.

2. DESIGN OF THE MATCHING NETWORKS

The PC-board material must guarantee low line losses at the required frequency and a high Q of the resonant circuit. Epoxy glassfibre boards can be used up to 1.3 GHz; at higher frequencies, PTFE glassfibre boards should be used. This article is not to discuss the width and length of stripline circuits here, since this has been described several times in (3) to (6). In our case, the striplines were calculated using a TI 59 calculation program developed by the author.

2.1. Matching Circuits

Optimum gain characteristics can only be achieved when the transistor is correctly matched. In the case of a complex impedance, matching conditions will be present when terminated with the conjugated complex value. In this case, the real impedance of the termination has the same value of that of the source; the reactive impedance of the termination should have the same value, but of opposite sign as that of the

source **(Figure 2)**. This means that an inductance is compensated for with a capacitance of the same reactive impedance, and vice versa.

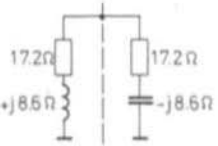


Fig. 2: Matching of the NE 578 35 input at 2 GHz

The matching of the reactive impedance can often be made directly at the transistor. In some cases, however, it must be carried out with the aid of a transformation link. This is to be discussed in detail. Since the real impedance of the transistor output will not coincide with the input of the following transistor, it is necessary for a transformation link to be provided. This is also the case between amplifier input or output and transistor input or output respectively.

There are several different methods of transformation possible. However, due to the simple calculation, only the λ/4 transformation is to be used here. This means that matching is only exhibited within a certain bandwidth on both sides of the required frequency.

2.1.1. Reactive Impedance Matching

If the reactive impedance component of the transistor impedance is positive, in other words inductive, it is possible for the matching to be made directly at the transistor with the aid of a capacitance. Due to the possible spread of the transistor specifications, a trimmer should be used, and this should be a ceramic tubular trimmer with a low minimum capacitance for UHF and SHF applications. A parallel circuit to ground is more suitable for alignment than a series circuit. For this reason, the transistor series impedance $Z_S = R_S \pm jX_S$ must be converted to an equivalent parallel impedance in order to calculate the required capacitance value. This is made with the aid of the following equations:

$$R_P = R_S + \frac{X_S^2}{R_S} \tag{3a}$$

$$X_P = \frac{R_S \times R_P}{\pm X_S} \tag{3b}$$

$$Z_P = R_P \pm jX_P \tag{3c}$$

Example:

$Z_{P\,in}$ of the NE 578 35 at 2 GHz is $21.5 \parallel + j\,43\,\Omega$. The required parallel capacitance is calculated according to equation 4 as follows:

$$C/\text{pF} = \frac{1}{2\pi f/\text{MHz} \times 10^6 \times X_P \times 10^{-12}} \tag{4}$$

C = 1,85 pF für 2 GHz.

The whole transformation process is given in **Figure 3**.

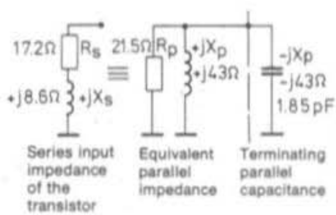


Fig. 3: Termination of the input reactive impedance of the NE 578 35 at 2 GHz

If the reactive impedance component of the transistor impedance is negative (capacitive), it is possible for matching to be made using an inductance directly at the transistor. Unfortunately variable inductances cannot be constructed in the SHF-range, which means that it is advisable to carry out a λ/4 transformation of the reactive impedance. The reactive impedance is transformed to an opposite value in a λ/4 line, which means that an inductance will be transformed into capacitance.

The transformed value can be calculated according to equation 5:

$$X_2 = \frac{-Z_0^2}{X_1} \tag{5}$$

where

Z_0 = Impedance of the transformation line.

If the transformation line is in series with the transistor impedance, the series-reactive impedance of the transistor is inserted in equation 5 for X_1. **Figure 4** shows this transformation process in the form of a circuit diagram.

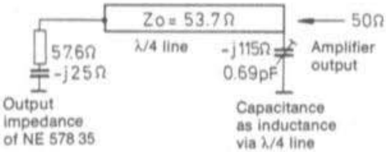


Fig. 4: Matching the output impedance of the NE 578 35 at 2 GHz and U_{CE} = 8 V / I_C = 10 mA into 50 Ω

One exception during the reactive impedance matching using variable capacitors is the output of the first amplifier stage. The reactive component of the output impedance of the transistors used is capacitive. Since this impedance must be transformed to the input impedance of the second stage, a fixed inductance is used at the collector of the first stage in order to provide a simple transformation path.

According to equation 6, an inductance can be realized by using a stripline circuit that is shorted at the end for radio frequencies. The impedance can be selected as required. Since the width of the stripline is mainly dependent on the impedance, a compromise must be made during the selection of the impedance:

The compensation line should be as narrow as possible in order not to have an effect on the λ/4 transformation line. On the other hand, in order to maintain the required impedance, it must be wide enough that the width can be accurately realized within ± 0.5 mm in the design drawing. In practice, a well-proved compromise is for the width of the compensating line to be w ≦ the width of the λ/4 transformation line.

In professional applications, such stubs are usually made in the form of two partial stubs in order to have as little effect as possible on the transformation line. This method was not used in the described amplifiers.

The following equation provides information for calculating a stripline as an inductance (end short-circuited !):

$$X_L = Z_0 \tan \Theta$$
$$[\Theta < 90^\circ \triangleq \lambda/4] \qquad (6)$$
$$\Theta\,[l/\lambda_0] = \frac{\arctan X_L/Z_0}{360}$$
$$[0, \text{ to } \lambda_0]$$

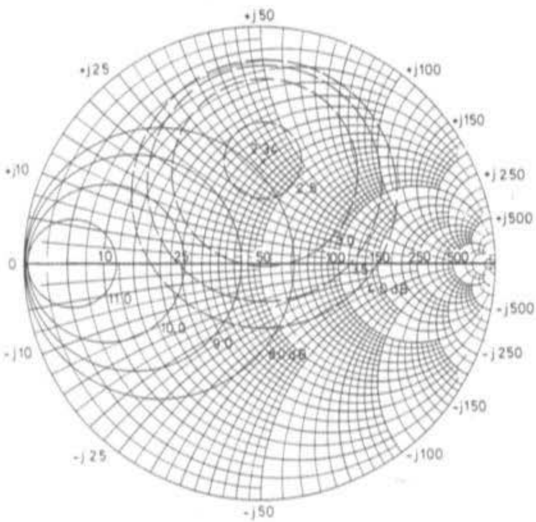


Fig. 5: Noise figure and gain curves of the NE 578 35 at 2 GHz and U_{CE} = 8 V and I_C = 3 mA

2.1.2. Real Impedance Matching

Only the real components of the complex impedances are used for calculation of the impedance of the λ/4 transformation line. The impedance results as given in equation 7 :

$$Z_0 = \sqrt{Z_1 \times Z_2} \qquad (7)$$

If a reactive element is connected in parallel with the transistor, it is necessary for the equivalent real, parallel impedance to be inserted as Z_1 or Z_2.

2.2. Noise Matching of the Input Stage

The lowest noise figure of an amplifier stage will not coincide with its optimum (power) matching, but with a certain degree of mismatch. The complex source-reflection coefficient Γ_0 is given by the manufacturers for alignment of an amplifier stage for minimum noise figure. The required source impedance can be obtained from this value with the aid of equation 2 or from the Smith diagram. The values for Γ_0 are not given in tabular form, but in the form of a Smith diagram.

Figure 5 shows such a diagram with circular contours. The gain curves (continuous lines) intersect the noise figure curves (with dashed lines). The given noise figure can be obtained within a noise figure curve. The matching gain value results from the dB-value of the intersected gain curve. For example: NF = 4 dB at G_p = 10 to 11 dB, or NF = 2.34 dB at G_p = 8 to 9 dB. It is thus possible to determine whether exaggerated values are given by the manufacturers of preamplifiers and converters by determining whether the given noise figure and gain values match those given for the transistor used.

This complicated method is only advisable when an exact calculation of the interactions within the transistor is to be taken into consideration. This method of calculation is part of the previously mentioned TI 59-program. The impedance obtained from s_{11} is used for a simplified noise matching.

In order to align the amplifier for minimum noise figure by measurement or by ear, the input network should be in the form of a variable Pi-filter that can be varied within wide limits. A Pi-filter is also advisable when no accurate s-parameters are available for the exact calculation, but only inter or extrapolated values. This is the case, for instance, for the 2.3 GHz version to be described later.

3. FURTHER CIRCUIT DETAILS

3.1. Operating Voltage Supply

The base and collector voltages must be fed to the transistors in such a manner that they do not affect the matching networks. This is achieved using extremely narrow λ/4 lines that are short-circuited for SHF at the ends. This short circuit is transformed by the λ/4 line to an infinite value so that it does not interfere with the matching lines at the transistor connections.

An impedance of 150 Ω was selected for the voltage supply lines. This means that they are only 0.1 to 0.4 mm wide.

The SHF-short circuit at the end of the lines is also achieved with the aid of λ/4 lines that are connected to the narrow voltage supply lines and are open at the other end. An impedance of 25 Ω was selected for these lines. Due to their relatively large area, these lines also represent capacitors and form an additional short circuit for such high frequencies.

The voltage is fed to the intersection of both lines via a small resistor.

3.2. Number of Amplifier Stages

According to well-known formulas, the overall noise figure of the receive system will have a low value if the noise figure of the preamplifier is considerably lower than that of the mixer, and when the preamplifier also possesses a high gain. For example, for a preamplifier with a noise figure of 3 dB and a gain of 10 dB in front of a mixer with a noise figure of 12 dB (for example a hybrid ring mixer), an overall noise figure of 5.4 dB would result. On the other hand, a preamplifier with the same noise figure but 20 dB gain in front of the same mixer would provide an overall noise figure of approximately 3.3 dB.

Since the gain of a transistor stage aligned for minimum noise figure is only listed as being 8 to 11 dB in the data sheets, it is necessary to use two amplifier stages in order to ensure that the low noise figure of the transistor is utilized to the full in the receive system. In this case, the first stage is aligned for minimum noise, and the second stage for maximum gain.

The large-signal behaviour has not been taken into consideration, and this is still not important on the SHF-amateur bands.

3.3. Operating Points

The operating points (U_{CE}, I_C) for minimum noise and maximum gain are given in the data sheets of the SHF-transistors. These voltages must be stabilized with respect to temperature and voltage fluctuations. The circuit given in (7) and (8) has been found to be very favorable for this and is used in the described amplifiers.

The stabilization circuit uses a voltage feedback to the base in conjunction with a base-constant current source. A voltage stabilization using a zener diode is used for stabilizing voltage fluctuations that can occur especially during portable operation. **Figure 6** shows the circuit diagram of this DC-circuit.

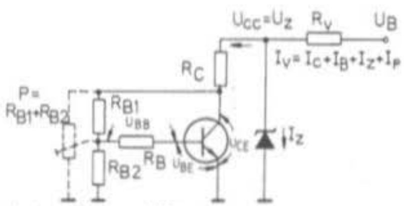


Fig. 6: Operating point adjustment and stabilization

The zener voltage should be approximately 1 V greater than the given collector-emitter voltage. The value of the dropper resistor R_d is selected so that twice the collector current flows through R_d at the nominal operating voltage of 12 V.

Example:

U_{CE} = 8 V; U_Z = 9.1 V; I_C = 10 mA;

U_B = 12 V.

$$R_d = \frac{2.9\ \mathrm{V}}{0.02\ \mathrm{A}} = 145\ \Omega \quad \text{(stand. value 150 } \Omega\text{)}$$

The fundamental considerations in the previously mentioned articles are not to be repeated here. However, the equations for Figure 6 are to be given:

$$R_B = \frac{h_{FE}\,(U_{BB} - U_{BE})}{I_C} \tag{8}$$

$$R_{B2} = \frac{h_{FE} \times U_{BB}}{5\, I_C} \tag{9}$$

$$R_{B1} = \frac{h_{FE}\,(U_{CE} - U_{BB})}{6\, I_C} \tag{10}$$

$$R_C = \frac{h_{FE}\,(U_{CC} - U_{CE})}{(h_{FE} + 6) \times I_C} \tag{11}$$

$U_{CC} = U_Z$;

U_{CE} and I_C from data sheet;

U_{BE} = 0.7 V for silicon transistors;

U_{BB} = 2 V for silicon transistors;

h_{FE} from the data sheet (assume h_{FE} = 100 if not listed).

For alignment of the operating point, the individual resistors R_{B1} and R_{B2} are replaced by a single trimmer potentiometer.

This temperature stabilization operates very satisfactorily: The collector current only varies by approximately ± 1 mA from the value selected at + 20°C throughout the temperature range of − 20°C and + 60°C !

The operating voltage may drop down to 10.5 V without altering the collector current. The voltage stabilization is especially advantageous since the maximum permissible collector-emitter voltage of the given NEC transistors only amounts to 11 and 12 V, respectively.

4. CONSTRUCTION OF THE AMPLIFIER

For the previously mentioned reasons, two-stage amplifiers were constructed for the 23 cm and 13 cm band equipped with tran-

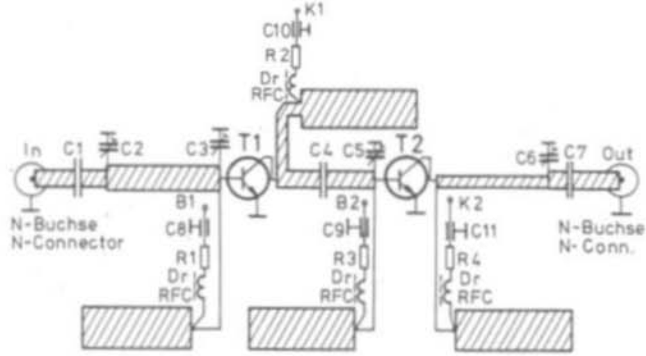


Fig. 7: SHF-circuit diagram for the described amplifiers

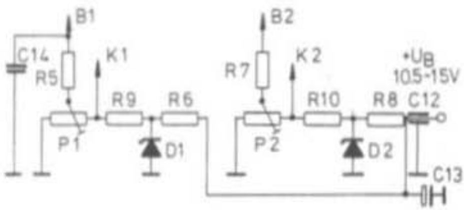


Fig. 8: Circuit diagram of the operating point adjustment and stabilization

sistor type NE 645 35 in the input and transistor NE 578 35 as second stage. The amplifier was constructed using a PTFE glassfibre board (1.57 mm thick). In addition to this, a cheaper version was developed for the 23 cm band equipped with two NE 578 35 using an epoxy glassfibre board, also 1.57 mm thick. **Figure 7 and 8** give the circuit diagram valid for all three amplifiers, separately for SHF and DC.

Components:

Band	T 1	T 2
23 cm inexpensive version	NE 578 35	NE 578 35
23 cm	NE 645 35	NE 578 35
13 cm	NE 645 35	NE 578 35

D 1, D 2:	Zener diode C 9 V 1
RFC :	Ferrite bead on resistor lead
C 1, C 4, C 7:	Chip capacitor 20-100 pF
C 2, C 3, C 5:	0.5 - 3 pF ceramic miniature tubular trimmer
(C 3 deleted for 23 cm amplifier)	
C 6:	0.5 - 6 pF ceramic miniature tubular trimmer
C 8, C 12:	Feedthrough capacitor 470 - 1000 pF
C 13:	10 µF / 25 V tantalum electrolytic
C 14:	47 nF ceramic capacitor
R 1 - R 4:	10 Ω (value uncritical)

Figures 9, 10 and 11 show the conductor side of the 23 cm preamplifier constructed on a PTFE board (DJ 6 PI 008) and of epoxy material (DJ 6 PI 009), as well as the 13 cm preamplifier on a PTFE board (DJ 6 PI 010). Each of the three versions is shown from both sides in the photographs given in **Figures 12 to 17.**

The ground surface of the boards is not etched. Holes of 3.2 mm diameter are drilled for the feedthrough capacitors C 8 - C 11 at four positions on the board marked with rings.

Band	R 5	R 6	R 7	R 8	R 9	R 10	P 1	P 2
23 cm inexpensive version	47 k	270	12 k	150	330	100	50 k	10 k
23 cm + 13 cm	18 k	220	12 k	150	150	100	25 k	10 k

Fig. 9:
PC-board of the 23 cm preamplifier using PTFE material

Fig. 12 and 13:
Photograph of the 23 cm preamplifier (PTFE) showing both sides

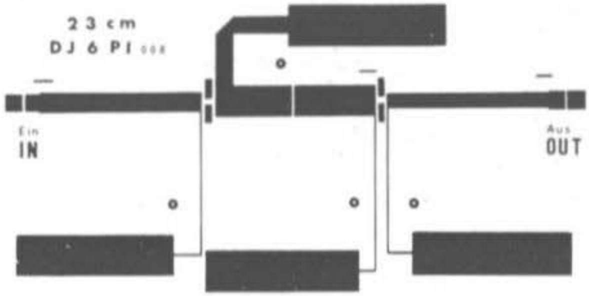


Fig. 10:
PC-board for the 23 cm preamplifier using epoxy material

Fig. 14 and 15:
Photograph of the 23 cm preamplifier (epoxy) from both sides

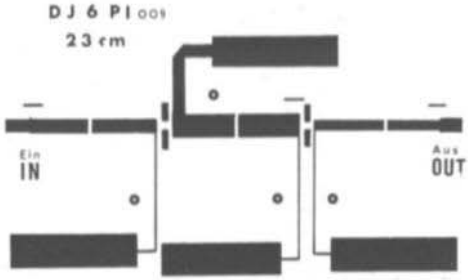
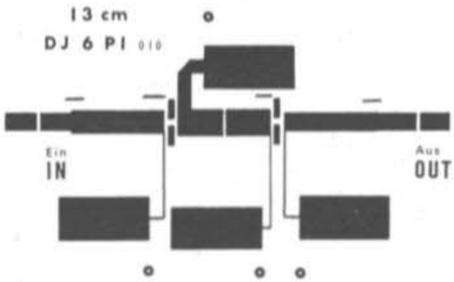


Fig. 11:
PC-board for the 13 cm preamplifier using PTFE material

Fig. 16 and 17:
Photograph of the 13 cm preamplifier



Some tendency to oscillation was observed with the first prototypes when the two emitter connections of the transistors were bent down and placed through 1 mm holes to the ground side of the board. For this reason, only small contact surfaces are provided for the emitter connections, which must be provided with low-inductive contact to the ground surface. This is achieved by sawing a small cut along the narrow side of this surface and placing an approximately 2 mm wide copper foil strip through this slot. The copper foil is then soldered to both sides of the board. **Figure 18** shows this in the form of a drawing.

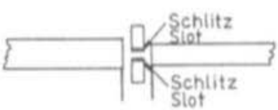


Fig. 18:
It is necessary for through-contacts to be made for the emitter connection surfaces

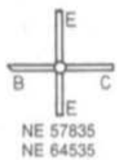
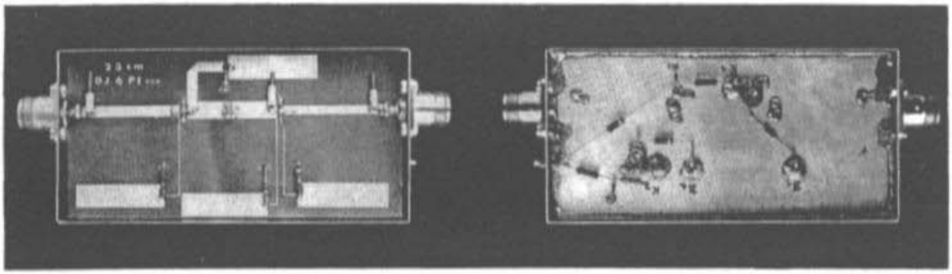
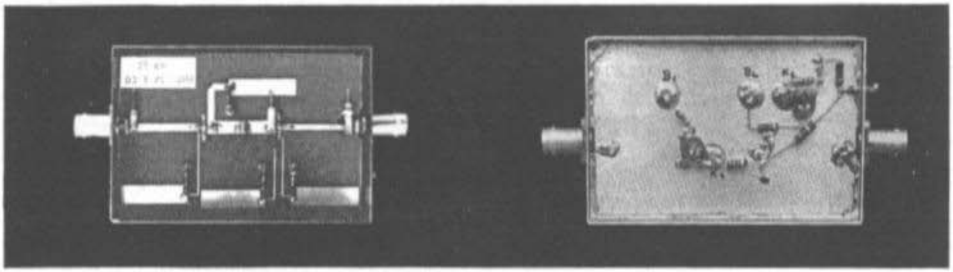
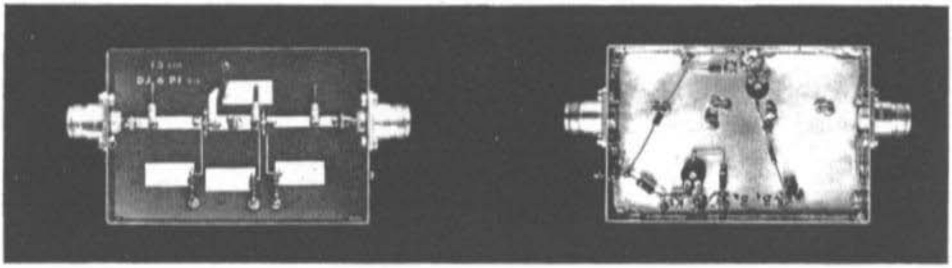


Fig. 19:
Connection diagram for the NEC transistors







After this, the transistors are soldered to the striplines without shortening the connection leads. In the case of the NEC transistors, the diagonally cut connection is the base (see **Figure 19**).

Due to their minute dimensions, the transistors can only be provided with a coded type designation: this is C_g (or another small letter) in the case of the NE 578 35, and Kz (or another small letter) for the NE 645 35.

The trapezoidal chip capacitors C 1, C 4, and C 7 are soldered into position vertically at the interruptions of the striplines. The ground connection of the ceramic tubular trimmers depends on the type used. In the case of the subminiature trimmers type RT 23 (Stettner), slots are cut into the board at the positions marked with lines, and the ground connection of the trimmer is then placed through the slot and soldered. In this manner, it is possible to achieve very short connections. The capacitance surface of the trimmers is then soldered to the striplines.

If trimmers without connection leads are to be used, a small bracket made from silver-plated wire should be bent around the

ground connection and the two ends of the wire placed through 1 mm diameter holes in the board and soldered both to the ground surface and the trimmer. The capacitance surface is soldered to the stripline.

The feedthrough capacitors are soldered to the ground surface. The low-value resistors for the base and collector voltages are soldered with short connections to the feedthrough capacitors. Ferrite beads are placed over the wire ends at the other side of the resistors after which they are soldered into position at the intersection of the wide and narrow lines.

The PC-board should be soldered into a case which allows the base and cover to be removed. It is, however, also sufficient when a screening panel constructed from thin brass, copper or tin plate is provided around the edge of the board. Holes of approximately 4 mm diameter should be provided for aligning the trimmers.

A small cutout must be provided at the input and output of the amplifier to fit the protruding insulation between the inner condcutor and flange of the BNC (N is better) connector. The inner conductor of the connector is directly soldered to the end of the striplines. This can be seen in the photographs.

The dropper resistors R 5 - R 10, the two trimmer potentiometers, the two zener diodes and the bypass capacitor C 14 are accommodated in the ground side of the case as shown in the photographs. The common 12 V connection is made via feedthrough capacitor C 12 in the screening panel of the case.

5. ALIGNMENT AND MEASURED VALUES

Before connecting the operating voltage for the first time, the trimmer resistors for the base voltage should be adjusted to their ground end stop. A mA-meter is now connected into the collector line of the transistor to be adjusted. The trimmers are now adjusted so that the following collector currents result:

Band	First stage	Second stage
23 cm inexpensive version	3 mA	10 mA
23 cm + 13 cm	7 mA	10 mA

A signal generator, beacon signal or a signal from another radio station is required for alignment of the trimmer capacitors. It is assumed that the mixer, oscillator and IF-amplifier have already been aligned correctly. In most cases, the signal will still be heard weakly inspite of the fact that the preamplifier has still not been aligned correctly. The trimmers are then aligned one after another in several processes for maximum gain. Finally, the input trimmer or trimmers should be aligned with the aid of a noise generator, or by ear, for best signal-to-noise ratio.

The bandwidth of these amplifiers is so great that the relatively wide amateur bands including any change of the 13 cm band are amplified without deterioration.

5.1. Measured Values

The gain was measured with a HP signal generator, multiplier and spectrum analyzer. The noise figure was measured with a AIL noise generator together with an interdigital converter as described by DC 0 DA and included preamplifiers subsequent to the preamplifiers to be measured. Since a narrow-band mixer is used, the measured noise figures are single-sideband noise figures of the preamplifier including a small contribution from the subsequent unit.

The PC-boards described here are designed for transistors type NE 578 35 and NE 645 35. If other transistor types are used, one must expect inferior values for gain and noise figure if the striplines are not recalculated.

If strong signals are present in the neighbouring frequency ranges, it may be necessary to provide a narrow-band, low-loss filter. Of course, the insertion loss of the filter will deteriorate the noise figure. Provision of a filter is also advisable when the noise component at the image frequency of a wideband mixer is to be eliminated. In

	Overall noise figure		Gain	
Band	Amplifier with simple calculation	Amplifier with exact calculation	Amplifier with simple calculation	Amplifier with exact calculation
23 cm inexpensive version	3.2 dB	2.5 dB	20 dB	26 dB
23 cm	2.5 dB	1.8 dB	23 dB	26 dB
13 cm	3.6 dB	2.5 dB	18 dB	21 dB

Table 2: Measured values of the three preamplifier modules

order to ensure that the noise figure of the preamplifier is not considerably deteriorated, the filter should be connected between preamplifier and mixer in this case. High-quality coaxial or interdigital filters should be used in all cases.

A similar preamplifier for the 9 cm band will be constructed using RT-duroid material as soon as activity warrants this. This will then be described in a future article.

6. COMPUTER PROGRAMS

As previously mentioned, programs have been developed for the miniature computer TI-59 which is programmed with magnetic cards. This allows the following to be computed:

- The output reflection coefficient s_{22}^* for noise matching of the first stage
- The corrected input and output reflection coefficients s_{11}^* and s_{22}^* for simultaneous input and output matching (both taking the forward and return transfer coefficients s_{21} and s_{12} into consideration)
- The transistor impedances from the reflection coefficients
- The stability behaviour and the calculated gain
- Matching networks (length and width of the striplines, dimensions of the reactive impedance elements)
- Resistance values for alignment of the operating points

The programs are available from the author at a cost of DM 100.—. They include three recorded TI 59 magnetic cards, the program list, input instructions, and a complete calculation example.

7. REFERENCES

(1) P. Shuch, WA 6 UAM: Microstripline Preamplifiers for 1296 MHz
ham radio magazine
April 1975, pages 12 - 27 and
October 1975, pages 42 - 46

(2) P. Shuch, WA 6 UAM: Solid-state Microwave Amplifier Design
ham radio magazine
October 1976, pages 40 - 47

(3) K. Hupfer: Striplines for VHF and UHF
VHF COMMUNICATIONS 3,
Edition 4/1971, pages 207 - 216

(4) W. Schuhmacher: Dimensioning of Microstripline Circuits
VHF COMMUNICATIONS 4,
Edition 3/1972, pages 130 - 143

(5) J. Fisk, W 1 HR: Simple Formula for Microstrip Impedance
ham radio magazine
December 1977, pages 72 - 73

(6) J. Fisk, HR:
Microstrip Transmission Line
ham radio magazine
January 1978, pages 28 - 37

(7) K. Richter: Design-DC-Stability Into Your Transistor Circuits
Microwaves, December 1973,
pages 40 - 46

(8) J. Fisk, W 1 HR:
Low-Noise 30 MHz Preamplifier
ham radio magazine
October 1978, pages 38 - 41